

UNITED STATES PATENT APPLICATION

For

HIGH-SPEED PHASE FREQUENCY DETECTION MODULE

Inventors:

Jonathan Cheah

Christopher Yong

Minjie Wu

Prepared by:

BLAKELY SOKOLOFF TAYLOR & ZAFMAN LLP
12400 Wilshire Boulevard
Los Angeles, CA 90025-1026
(408) 720-8300

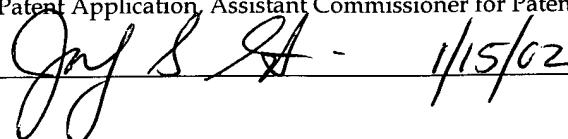
Attorney Docket No.: 004939.P023

"Express Mail" mailing label number: EL867651107US

Date of Deposit: January 15, 2002

I hereby certify that I am causing this paper or fee to be deposited with the United States Postal Service "Express Mail Post Office to Addressee" service on the date indicated above and that this paper or fee has been addressed to Box Patent Application, Assistant Commissioner for Patents, Washington, D. C. 20231-0001

Judy L. Steinkraus



Judy L. Steinkraus - 1/15/02

HIGH-SPEED PHASE FREQUENCY DETECTION MODULE

FIELD OF THE INVENTION

[0001] The present invention relates generally to the field of high frequency data transmission systems and, more specifically, to phase-lock-loops, synthesizers, and high-speed phase frequency detection devices.

BACKGROUND OF THE INVENTION

[0002] As wireless communication devices increase in number and complexity, there is an ever increasing need for wireless communication devices and systems that can function at very high frequencies. To this end, high-speed phase-frequency detection devices that can handle higher frequencies are needed.

[0003] Typically, phase-frequency detection devices are used in phase-lock-loops. A phase-lock-loop compares the phase and/or frequency of a generated clock signal (or more often a divided version of it) to that of a pure reference clock signal of known accurate frequency and adjusts the frequency and phase of the generated clock signal until it is "locked" in both frequency and phase to the reference clock signal. Thus, a clock signal of a high frequency that is a multiple of the reference frequency is established, which becomes the base for high-frequency wireless communication systems.

[0004] Figure 1A is an illustration of a typical phase-lock loop 100. It includes a phase-frequency detector ("PFD") 102, a charge pump 104, a voltage controlled

oscillator (“VCO”) 106, a loop filter 108, and most often a divider 110. The PFD 102 compares the phase and frequency of a reference clock signal (Reference Clock) to a feedback clock signal (Feedback Clock). The Feedback Clock signal is generated by the divider 110, which divides down from the VCO 106 output (Output Clock). If the Feedback Clock and the Reference Clock are out of phase, the PFD 102 will generate corrective voltage signals (V_{up} or V_{down}) to trigger the charge pump 104 to either source current into (I_{up}) or sink current from (I_{down}) the loop filter 108, hence increase or decrease the control voltage (V_{cntl}) going into VCO 106. The control voltage drives the frequency of the clock signal higher or lower, until the clock signal and the reference signal are matched in phase and frequency, or “locked.” When the loop locks, the charge pump 104 should stop sourcing or sinking current, thus leaving the control voltage relatively constant.

[0005] However, even when locked, the PFD 102 may experience jitter, or small noise error, in what is known as the “dead zone,” depicted in **Figure 1B**. Jitter as a result of dead zone occurs when the Feedback Clock signal varies slightly from the Reference Clock signal and very small phase differences develop ($\Delta\phi$). When the phase difference is slight enough, finite rise and fall times of the PFD pulse signals, caused by circuit-element capacitance, may not find enough time to reach a good logic high level, hence failing to turn on/off the charge pump

current switches. This leads to random variation in the VCO 106 output, or jitter.

[0006] One solution to eliminate jitter in the dead zone is to generate overlapping reset pulses, with wide enough pulse widths, to turn on/off the charge pump switches, also shown in **Figure 1B**.

[0007] **Figure 1C** shows the input-output characteristic of the ideal PFD, PFD with dead zone, and PFD with a reset pulse. Generally, when overlapping reset pulses are too wide, more information of phase relationship is lost, which affects the performance of the phase-lock-loop system. Thus, the reset pulse has to remain narrow, compared to the period of the clock signals, to achieve good loop dynamics.

[0008] Many communication systems today are operated at very high frequencies, such as the 5.18GHz to 5.805GHz frequency spectrum used for the 802.11a standard. Unfortunately, current phase-lock-loop devices, such as PFDs (see **Figure 1D**) and charge pumps (see **Figure 1E**), are too slow and thus cannot create corrective pulses and control voltages for very high frequencies. In addition, current charge pumps, like the one shown in **Figure 1E**, require complicated feedback and replica circuit to function properly. Such excessive circuitry add complexity to the circuit design and cause instability when used at high frequencies.

BRIEF DESCRIPTION OF THE DRAWINGS

[0009] The present invention is illustrated by way of example and not limited by the figures of the accompanying drawings in which like references indicate similar elements and in which:

[0010] **Figure 1A** is an illustration of a conventional phase-lock loop.

[0011] **Figures 1B-C** are a graphical illustration describing deadzone, the corrective measures and its consequence.

[0012] **Figure 1D** is an illustration of a conventional PFD.

[0013] **Figure 1E** is an illustration of a conventional charge-pump.

[0014] **Figure 2** is an illustration of one embodiment of a high-speed phase-frequency detection module 200 configured according to the present invention.

[0015] **Figure 3** is an illustration of one embodiment of a high-speed PFD 300 configured according to the present invention.

[0016] **Figure 4** is an illustration of one embodiment of a high-speed, edge-triggered asynchronous reset TSPC D-flip-flop 400 configured according to the present invention.

[0017] **Figure 5** is an illustration of one embodiment of a high-speed charge-pump 500, according to the present invention.

[0018] **Figure 6** is an illustration of one embodiment of a bias circuit 600 according to the present invention.

[0019] **Figure 7** is an illustration of one embodiment of a frequency synthesizer

700 configured according to the present invention.

SUMMARY OF THE INVENTION

[0020] A high-speed phase-frequency detection module is described to function at very high frequencies and to produce very low jitter. In one embodiment, high-speed phase-frequency detection module includes a PFD with edge-triggered asynchronous-reset true-single-phase-clocking ("TSPC") D flip-flops that have very short CLK to Q and Reset to Q time delays. In one embodiment, the high-speed phase-frequency detection module includes a charge pump, without replica or feedback, that can respond to the very narrow pulses from the PFD and produce output voltages with small ripple, thus leading to low VCO output jitter. Other features of the present invention will be apparent from the accompanying drawings and from the detailed description that follows.

DETAILED DESCRIPTION OF THE INVENTION

Described herein is a high-speed phase-frequency detection module. In the following description numerous specific details are set forth in order to provide a thorough understanding of the present invention. One of ordinary skill in the art, however, will appreciate that these specific details are not necessary to practice the present invention. While certain exemplary embodiments have been described and shown in the accompanying drawings, it is to be understood that such embodiments are merely illustrative and not restrictive of the current invention, and that this invention is not restricted to the specific constructions

and arrangements shown and described since modifications may occur to those ordinarily skilled in the art.

[0021] **Figure 2** is a diagram of a high-speed phase-frequency detection module 200 configured according to the present invention. The high-speed phase-frequency detection module 200 includes a high-speed PFD 202 and a fast-reacting charge pump 204. The high-speed PFD 202 includes a first input to receive a very high frequency Reference Clock signal and a second input to receive a Feedback Clock signal. The high-speed PFD 202 is configured to compare the phase and/or frequency difference of the Feedback Clock signal and the Reference Clock signal and produce voltage pulses on its outputs (UP/XUP and DOWN/XDOWN). The voltage pulses are received by the high-speed charge pump 240, which is configured to provide corrective currents (PUMP up/down). Further details of the high-speed PFD 202 are discussed in conjunction with **Figure 3**. Further details of the charge pump are discussed in conjunction with **Figure 5**.

[0022] **Figure 3** is a diagram of one embodiment of a high-speed PFD 300 configured according to the present invention. It includes two edge-triggered asynchronous-reset TSPC D flip-flops ("TSPC DFF") 322 and 324, and an OR gate 326. TSPC DFFs 322 and 324 are configured to provide very short CLK to Q and Reset to Q delays, as low as sub-100ps. An exemplary embodiment of the TSPC DFFs 322 and 324 is described in conjunction with **Figure 4**.

[0023] The first TSPC DFF 322 has a first input 330 to receive the Reference Clock signal, a second input 332 to receive a reset signal (R), and outputs 334 and 336 to produce output signals (UP and XUP, where XUP is the inverse of UP). The second TSPC DFF 324 has a first input 340 to receive the Feedback Clock signal, a second input 342 to receive a reset signal (R), and outputs 344 and 346 to produce output signals (DOWN and XDOWN, where XDOWN is the inverse of DOWN).

[0024] The TSPC DFFs 322 and 324 are configured to compare the Reference Clock signal and the Feedback Clock signal to determine if they match phase or frequency. If the phase of the Reference Clock signal leads that of the Feedback Clock signal, then the first TSPC DFF 322 will produce a first voltage signal (i.e. UP/XUP), and send it to the charge pump. If the phase of the Reference Clock signal lags that of the Feedback Clock signal, then the second TSPC DFF 324 will produce a second voltage signal (i.e. DOWN/XDOWN), and send it to the charge-pump. *324*

[0025] The OR gate 326 includes a first input 352 to receive the data output signal XUP from TSPC DFF 322, a second input 354 to receive the output signal XDOWN from TSPC DFF 324, and a reset output 356 to produce a reset output signal (R).

[0026] In one embodiment, all of the TSPC DFF outputs (UP/XUP and DOWN/XDOWN) are brought out to interface with charge pump without additional inverting logic. For the high-speed PFD 300 to function at high

speeds, it is advantageous to not have inverting logic because it is too slow. At very high frequencies the pulse widths of the TSPC DFF output signals (UP/XUP and DOWN/XDOWN) are very narrow, and any additional inverter delay can slow them down. Additionally, any digital logic processing can slow down the output signals. Hence, in one embodiment, no digital processing logic operates on the TSPC DFF output signals between the high-speed PFD 300 and the subsequent charge-pump. Small static CMOS transistors are used to perform the inverted logic within the TSPC DFFs 322 and 324. The CMOS logic is very fast and power efficient. The high-speed PFD 300 can produce reset pulses with widths as small as 200ps.

[0027] **Figure 4** is an illustration of an exemplary high-speed, edge-triggered asynchronous reset TSPC DFF 400 configured according to the present invention. TSPC DFF 400 includes a P-precharged Domino logic gate (“p-block”) 402, followed by an n-precharged Domino logic gate (“n-block”) 404, followed by an inverting buffer 406. The p-block does not include an inverting clock latch, but the n-block does. The p-block and n-block are clocked, hence only the input and output of the clocked latch of n-block requires resetting. Consequently, TSPC DFF 400 can provide very short CLK to Q and Reset to Q delay, as low as sub-100ps.

[0028] **Figure 5** is a diagram of one embodiment of a high-speed charge-pump 500, according to the present invention. Referring to **Figure 5**, the charge-pump

500 includes an upper current branch 502 to source a positive current pulse. In one embodiment, the upper current branch consists of a plurality of PMOS transistors 511 - 514. The charge-pump 500 also includes a lower current branch 504 to sink a negative current pulse. In one embodiment, the lower current branch 504 consists of a plurality of NMOS transistors 521 - 524.

[0029] The lower current branch 504 is connected to the upper branch 502 at two nodes, A and B. The output currents can be summed together by shorting at node B. An amplifier 506 is connected between the two nodes A and B.

[0030] The charge-pump 500 further includes a bias circuit 510, with current sources that are continuously on, coupled to the upper and lower current branches 502 and 504. An advantage of charge-pump 500 is that it does not utilize a feedback circuit or replica circuit, thus allowing the charge-pump 500. With no feedback or replica circuits, charge-pump 500 is simple in design and able to stabilize very quickly.

[0031] Transistors 511, 512, 521 and 522 are switching transistors that steer the current from one branch 502 to the other 504. When UP and DOWN are both low, current flows through the left sides of the branches 502 and 504, depicted by current path 1. When UP and DOWN are both high, current flows through the right sides of the branches 502 and 504, depicted by current path 2. In both cases, the net current flowing through the output is zero. When UP is high, and DOWN is low, current will flow from a current source in the bias circuit 510, through

transistors 521 and 523 to charge up the loop filter 550, while the other current path is from the unity gain buffer output through transistor 512 and 514 to sink current, depicted by current path 3. The operation when DOWN is high and UP is low is just the opposite, depicted by current path 4. Transistors 513, 514, 523, and 524 are configured to always be on and to minimize and eliminate charge injection effects, hence reducing the voltage ripples in the charge-pump 500 voltage (Vctl), thus leading to much less VCO output jitter. The op-amp 506 is used as a unity gain buffer and is configured to maintain the voltage potential of Vx and Vy to be very close to Vctl (by $2 \times V_{dsat}$) so that charge sharing effects are minimized.

[0032] **Figure 6** is an illustration of a bias circuit 600 according to the present invention. The bias circuit 600 is implemented using a standard cascode current mirror. An advantage of the bias circuit 600 is that it has very good accuracy and wide dynamic range, and hence eliminates the need for replica or feedback circuitry. This simplifies the design, and is faster because the closed loops of replica and feedback circuit require time to stabilize, and thus should not be used for very high frequencies.

[0033] **Figure 7** is an illustration of one embodiment of a frequency synthesizer 700 configured according to the present invention. It includes a high-speed PFD 702, a high-speed charge pump 704, a loop filter 706, a VCO 708, a prescaler device 712, and a programmable pulse swallow counter device 714. In one

embodiment, the high-speed PFD 702 and the high-speed charge pump 704 are configured according to any of the embodiments of the present invention described herein. Consequently, the frequency of the Reference Clock can be very high. In one embodiment, the Reference Clock frequency is 120MHz and the output frequency range is from 5G to 6G. Because of the advantages of the high-speed PFD 702 and the high-speed charge pump 704, the frequency synthesizer 700 has very little noise, is fast, and is simple in design.